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Specification and Drawings, as originally filed, with Application for Patent Serial No: **2,281,236**, on September 1, 1999, by **TAJINDER MANKU**, for "Direct Conversion RF Schemes Using A Virtually Generated Local Oscillator".

PRIORITY DOCUMENT
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ABSTRACT

This inventions has applications in RF receivers and transmitters. The patent describes a method of removing the LO-leakage problem associated with direct RF receiver or transmitter topologies. This problem is solved by generating a virtual LO signal within the RF signal path. The LO signal is constructed by using signals that do not contain a significant amount of power (or no power at all) at the LO frequency required. Any errors added to the wanted LO in the new topology are minimized using a closed loop.

I. Introduction

It is known in the field that direct RF down conversion or RF up conversion schemes have a serious problem with the local oscillator (LO) leaking into the RF path. The problem stems from the fact that in direct conversion schemes the LO frequency is equal to the wanted RF frequency. Therefor when the LO leaks into the RF path, its power is placed directly in the RF signal band. This causes the information stored in the RF signal band to be modified or/and distort.

The frequency conversion topology we are proposing uses a completely new idea than all existing topologies. In our topology, two well-defined signal are used to remove the LO leakage found in a direct conversion receiver. Furthermore, the signals provide a means to select the RF channel. Because of the nature of the two defined signals, this concept is new.

The invention is similar to direct conversion, but provides two fundamental advantages:

- Zero LO leakage into the RF band
- reduces the 1/f noise problems

The topology provides two basic advantages over a superheterodyne topology:

- Removes the second LO in a superheterodyne system
- Removes the requirement for RF and IF image rejection filters

In superheterodyne receivers more than one frequency conversion are preformed.

The new structure requires circuit integration in order to make it work well. Furthermore, it does not require both gain matching and phase (delay) matching to reject images as in an image rejection mixers – i.e. only one is required.

The new topology has numerous applications. Some applications include cellular, wireless phones, pagers, two way pagers, local area wireless networks, wireless models, wireless email, etc, and various satellite applications.

The new topology can accommodate RF bands with varying bandwidths.

II. Receiver Architecture

In this section we shall systematically derive the new receiver architecture. A generalization of the topology within the transmitter section will be discussed in sections IV and V. In the process of arriving at the new architecture, we shall review one very important concept which is used. This concept is the *spread spectrum* technique.

In Fig. 1 we have illustrated a simplified representation of how spread spectrum techniques are used in today's *code-division-multiple-access* (CDMA) wireless systems. The signals $g_i(t)$ are the so-called "spreading signals". These signals represent pseudonoise (*PN*) having a chip rate of $f_{cr}=1/T_{cr}$ and have the two states $+1$ and -1. By multiplying the data $d_i(t)$ by $g_i(t)$, the bandwidth of the data signal spreads. In Fig. 1, N transmitted channels are shown all of which are centered on the same RF carrier. In order to reconstruct the data stream $d_i(t)$, the RF signal is converted back to base band using some conversion technique (this is not important in this description) and is multiplied by the PN sequence of the i^{th} data stream, $g_i(t)$. This sequence is typically hard coded into the receiver terminal as a random seed generator. Along with the data $d_i(t)$, there is an additive noise arising from all the other "spreaded" channels (denoted as $n(t)$). One important advantage of a spread spectrum system is its ability to reject large in band interferes. As we shall show, the new architecture uses this basic spread spectrum idea to solve all the problems associated with the conventional direct conversion receivers.

The basic building block of the new architecture is shown in Fig. 2. The structure consists of two mixers (multipliers). The input to the system is denoted by *x(t)*. The term x(t) contains the wanted RF signal/channel/band. The goal of the receiver is to move the desired RF signal down to base band (i.e. centered on DC). The first mixer, M1 multiplies the signal x(t) with a periodic signal have two states +1 and –1 (denoted as $S_o(t)$) and by a LO signal which contains a large amount of power at the RF frequency; a generalization of the functional behavior of So will be given later in this document. The term LO represents the local oscillator tuned to the RF frequency of the desired channel. Under ideal conditions, the output M1 is given by the expression,

$$P_1(t) = S_o(t)x(t)LO \tag{1}$$

Note that the term $x(t)LO$ represents the base band spectrum of the desired RF channel. The above equation represents the spreading of the base band signal over the infinity bandwidth of $S_o(t)$. In this spectrum, all bands located kf_{So} away from the desired RF channel are aliased together, where *k=1,3,5,...* and f_{So} is the frequency of $S_o(t)$; see Fig. 3. One may initially think that it is counter intuitive to intentionally imposing this type of aliasing, however this is done all the time in spread spectrum type systems. The second mixer, M2 multiples (1) by $S_o(t)$, giving an output of,

$$\begin{aligned} P_2(t) &= S_o^2(t)x(t)\cos\omega_{RF}t \\ &= x(t)\cos\omega_{RF}t \end{aligned} \tag{2}$$

Under ideal conditions, $P_2(t)$ is simply the base band signal of the wanted RF channel. The above step is similar to the de-spreading step that occurs in spread spectrum systems. Consequently, the goal of M2 is to de-alias all the aliasing that occurred in the previous mixer. The question is why alias everything and then de-alias it? One obvious advantage is that any large interferes after M1 will less likely compress or block any gain stages between the mixers M1 and M2. The other advantages can only be seen if we incorporate all the various problems associated with a conventional direct conversion receiver. The main problems with a direct conversion receiver are the DC offsets due to

internal offsets, LO-RF leakage, RF-LO leakage, RF-base band leakage, and 1/f noise. In the following sections we have illustrated how these problems can be solved using the topology shown in Fig. 2.

Analysis of DC offsets, and LO-RF, RF-LO, and RF-base band Leakage

By including all the various DC offsets and signal-to-signal leakage terms, the output of mixer 1 is given by the expression,

$$P_1(t) = S_o(t)x(t)LO + \gamma_{23}S_o^2(t)LO^2 + \gamma_{53}S_o(t)LO^2 + \gamma_{31}x(t) + \gamma_{32}^2 x^2(t) + DC_{m1} \tag{3}$$

where γ_{ij} is the leakage of the signals at node i to node j (see Fig. 2), and DC_{m1} is the internal DC offset of M1. The term $S_o(t)x(t)LO$ represents the wanted term. All other terms in (3) are unwanted. The second term $\gamma_{23}S_o^2(t)LO^2$ represents the leakage of node 2 to node 3. This term encompasses two physical leakage mechanisms. These include:

1. Capacitive coupling within the circuit topology of the mixer
2. IC substrate coupling

In both of these cases, the coupling in non-deterministic; i.e. it changes with the physical surroundings of the receiver. When a signal from node 2 couples to node 3, a proportion of that signal will radiate into air via the antenna. This occurs because there is a finite reverse gain of the receiver. Depending on the surroundings, some proportion of this energy is reflected back into the receiving antenna where it is amplified, thus contributes to the term γ_{23}. The third term, $\gamma_{53}S_o(t)LO^2$ represents the leakage of the RF tuned frequency of the system to node 3. This leakage can occur in several ways depending on the architecture of the frequency synthesizer. These include:

1. Radiation of the RF tuned frequency into the receiving antenna
2. IC substrate coupling

The leakage associated with γ_{53} can not be ignored. However, through careful design, one could avoid synthesizing LO and directly synthesize

$S_o(t)LO$; this will be discussed later. The fourth term, $\gamma_{31}x(t)$ is related to the RF leakage to node 3. This term, in general could have a large DC term which arises from previous stages of the receiver. For example, nonlinear quadratic terms in the low noise amplifier could lead to DC terms. This is especial important when large interferes are present. The term $\gamma_{32}x^2(t)$ arises from the RF terms leaking into node 2. This term produces a DC term. This term is especially important in cases where *x(t)* contains large valued interferes around the wanted RF channel. The term $\gamma_{32}x^2(t)$ can be significantly reduced if differential circuits are employed. The last term, DC_{m1} denotes the internal DC offset of the mix structure itself. For a conventional direct conversion receiver the terms, γ_{23}, γ_{53}, γ_{31}, γ_{32}, and DC_{m1} all produce DC terms into the base band signal. This tends to reduce the sensitivity of the receiver structure. The output of M2 can be written as,

$$\begin{aligned} P_2(t) &= x(t)LO + \gamma_{23}S_o(t)LO^2 + \gamma_{53}LO^2 + \gamma_{31}S_o(t)x(t) \\ &+ \gamma_{32}S_o(t)x^2(t) + S_o(t)DC_{m1} \end{aligned} \tag{4}$$

or,

$$P_2(t) = x(t)LO + \gamma_{53}LO^2 + y_oS_o(t) \tag{5}$$

where,

$$y_o = \gamma_{23}\cos^2\omega_{RF}t + \gamma_{31}x(t) + \gamma_{32}x^2(t) + DC_{m1} \tag{6}$$

From (5) and (6) we can see that all except one DC offset term is pushed away from base band; Fig. 4 depicts the spectrum of equation (5). The DC offsets (absorbed in the term y_o) are translated to the frequencies kf_{So}. The only DC offset that remains arises from the radiation or substrate leakage term γ_{53}. As we shall describe later how this leakage term can be nulled to zero.

From the results above the frequency of $S_o(t)$ has to be greater than the bandwidth of the RF channel. If the frequency of $S_o(t)$ is made smaller than

this bandwidth, the DC offset terms (i.e. y_o) would fall directly within the RF channel causing a reduction in sensitivity.

In the analysis above, we have assumed $S_o(t)S_o(t)=+1$. However, in reality this is impossible to realize. By assuming $S_o(t)S_o(t)=+1$, we are assuming that both S_o's are perfectly in phase, and both $S_o(t)$'s have a rise and fall time of zero seconds. Under the most general of cases,

$$S_o(t)S_o(t-\tau)=1+\varepsilon(t,\tau) \tag{7}$$

where $\varepsilon(t,\tau)$ is an error function which has a period of $1/(2f_{So})$, and τ denotes the delay between the S_o's; see Fig. 5. The error function $\varepsilon(t,\tau)$ accounts for the fact the two S_o's are not perfectly in phase (denoted as CASE I) and the S_o's have a non-zero rise time (denoted as CASE II).

CASE I: If the two S_o's are delayed by τ the term $S_o(t)S_o(t-\tau)$ can be expressed as DC level of +1, plus a rectangular wave with duty cycle $100\times 2f_{so}\tau$ alternating between 0 and –2 (i.e. $\varepsilon(t,\tau)$); see Fig. 5. If the duty cycles is very small, $\varepsilon(t,\tau)$ will have spectral components sitting at frequencies $0,\pm 2f_{so},\pm 4f_{so},\ldots$ with magnitudes $\le 4f_{so}\tau$. By including the error function into equation (2), one obtains the results,

$$\begin{aligned} P_2(t) &= x(t)S_0(t)S_o(t-\tau)LO \\ &= x(t)LO+\varepsilon(t,\tau)x(t)LO \end{aligned} \tag{8}$$

The term $\varepsilon(t,\tau)x(t)\cos\omega_{RF}t$ absorbs the aliasing/imaging of the all channels located away from the desired RF channel by $2f_{so},4f_{so},\ldots$ into the RF channel itself. Though $2f_{so}\tau$ can be made small, it will never equal zero. Consequently, aliasing/imaging is enviable; see Fig. 6.

CASE II: To realize S_o within a hardware environment is impossible since the frequency spectrum of S_o expands from $-\infty$ to $+\infty$. In reality what we get is a "low pass filtered" version of S_o; this is illustrated in Fig. 7. Also shown in Fig. 7 is $LPF(S_o) \times LPF(S_o)$. As before we can write,

$$LPF(S_o) \times LPF(S_o) = 1 + \varepsilon(t,\tau) \qquad (9)$$

where $\varepsilon(t)$ has a period of $1/2f_{so}$. To first order, the magnitude of $\varepsilon(t,\tau)$ at the various harmonics is given by its rise time divided by its period. By including this error function into equation (2) we get,

$$P_2(t) = x(t)LO + \varepsilon(t)x(t)LO \qquad (10)$$

As in equation (10), the term $\varepsilon(t)x(t)LO$ gives rise to aliasing/imaging.

In both CASE I and II, the power contained within $\varepsilon(t)x(t)\cos\omega_{RF}t$ determines the amount of aliasing power brought back to base band. If ε is assumed to take the form,

$$\varepsilon = \begin{cases} -2 & nT < t < nT + \tau \\ 0 & nT + \tau < t < (n+1)T \end{cases} \qquad (11)$$

where $T=1/(2f_{so})$. In assuming the above form for ε, we are assuming the worst case scenario in terms of aliasing. The spectral components for ε are equal to,

$$\varepsilon = \frac{2}{n\pi}\sin\frac{n\pi\tau}{T} \qquad (12)$$

where $n = \pm1, \pm2, \ldots$. The relative amount of aliased power that would be brought to base band (i.e. the wanted band) is equal to

$$P_{aliased} = \frac{8}{\pi^2}\sum_{n=1,2\ldots}\frac{1}{n^2}\sin^2\frac{n\pi\tau}{T} \qquad (13)$$

Here we have made the assumption that the aliased noise by each harmonic is un-correlated. Equation (13) is finite in value, but converses at rate that depends on the ratio τ/T.

Though the above architecture may work for some RF applications it still requires a significant amount of RF filtering prior to entering the system to reduce the amount of energy in x(t). In the section below we shall present an error correction scheme for reducing the amount of aliasing.

Generation of Modulated LO signal

From equation (5) we see that if LO is synthesized directly there will always be a non-zero leakage into the RF band via the term γ_{53}. This is mainly due the fact the term S_oLO was initial said to be generated by multiplying S_o by the LO signal. However, the term S_oLO can be generated without generating a tone LO. This can be done directly in a phase lock loop or by using various other digital means such as direct digital synthesis. The basic criteria is to generate S_oLO without generating any power (or a relatively small amount of power[1]) at the frequency ω_{RF}. An example of generating S_oLO is illustrated in Fig. 8. In this figure a $2\omega_{RF}$ tone is used to generate S_oLO_I and S_oLO_Q without generating a frequency at ω_{RF}. Here S_o is denoted by the symbol PN and LO_I and LO_Q are the quaduature components of the LO signal. In this structure a timing corrected So is generated and is applied to the second mixer. The reasons for generating S_oLO_I and S_oLO_Q is so that the input x(t) can be decomposed into quaduature. This is useful for modulation schemes that are inherently in quaduature such as MSK, QPSK, PSK, GMSK, etc.

Timing correction

The output of the two mixers can be written in the form $x(t)LO(1+\varepsilon_{LO}(t,\tau))$ where $\varepsilon_{LO}(t,\tau)$ denotes the error in generating the correct LO signal. In Fig. 9 we have illustrated examples of this error under two different cases. The term

$x(t)LO\varepsilon_{LO}(t,\tau)$ contains two terms at base band: (I) aliasing power, (ii) power of the wanted signal, but at a reduced power which is on the order of $(\tau/T)^2$. Therefore the power at baseband (denoted by P_M) can be decomposed in to ~~three components: (I) the power of the wanted signal, P_w, (ii) the power of the~~ aliasing terms, P_a, and (iii) the power of the wanted signal arising from the term, $P_{w\varepsilon}$ (this power can either be positive or negative). Therefore,

$$P_M = P_w + P_{w\varepsilon}(\tau) + P_a(\tau) \tag{14}$$

Note that $P_{w\varepsilon}$ and P_a are a function of τ. If the power, P_m is measured and τ is adjusted, one can reduce the terms $P_{w\varepsilon}$ and P_a to zero. Mathematically this can be done if the slope of P_m with the delay τ is set to zero; i.e.,

$$\frac{dP_M}{d\tau} = 0 \tag{15}$$

A system diagram of this procedure is illustrated in Fig. 10. The power measurement scheme and the element blocks required to check if $\frac{dP_M}{d\tau} = 0$, can be implemented within a digital processing unit (DSP). The control signal instructing S_o to change it's delay is then applied to a controlled delay or rising/falling unit. Also illustrated in Fig.10 is a visual representation of the power measured versus delay. In this plot, we see that there is an optimum point at which $\frac{dP_M}{d\tau} = 0$. The basic criteria of this scheme is that the power measurement is made over a time, T_p shorter than the average time it takes for the power level of the wanted band width to change with time (this time is denoted by T_{pw}); i.e.

[1] "small" meaning that the amount of power generated at the RF carrier frequency is small enough that it

$$T_{pw} >> T_p \tag{16}$$

It is important to mention that the delay correction can be continuously correcting the delay, OR it may also be correcting it periodically, OR in some random sequence, OR only during the powering up of the RF receiver section.

Example of RF receiver with Timing corrected So signal[2]

In Fig. 11[3] illustrates a system diagram of the timing corrected S_o scheme. The items within the system diagram are:

M1I - the first input mixer multiplies the RF signal by LOI*PN where PNI is a function that varies from +A to -A where A is any number[4]. The sequence LOI*PN is generated using the techniques prescribed in the document. The design of this mixer[5] depends on the system specifications of the RF system.

M2I - The second mixer multiplies the output of M1I by PNI. PNI has been corrected for delay using the prescribed techniques. The basic idea here is that LOI*PN*PNI is equal to LOI. The design of this mixer[6] depends on the system specifications of the input RF signal. Though it has not been shown in the system diagram, a filter could be placed between M1I and M2I. However, in most applications it may not be necessary. The wanted signal at the output of M2I is lying at base-band (i.e. centered around DC).

LPF1I - This is a low pass filter. It is used to reduce the amount of out of band power, which may cause the following elements to compress in gain or distort the wanted signal. The design of this LPF depends on the bandwidth of the wanted signal.

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does not significantly degrade the performance of the RF receiver.

[3] Though the figure implying various elements are implemented in analog form they can be implemented in digital form.

[4] PN can be pseuso-random or a fixed periodic function - in this document we have assumed PN is equal to So and is a square wave with a 50% duties cycle and A=1.

[5] Design specifications include conversion gain, noise figure, and linearity.

[6] See footnote 3

SUM1I - The summing element is used to remove any DC offsets. Any known art can be used here.

LPF2I - This filter provides further filtering of the base-band signal. The design of this filter depends on the system specifications and system design trade offs.

GAIN-I - These elements provide a significant amount of gain to the base band signal. The design of this gain unit depends on the system specifications and system design trade offs.

LPF3I - This filter is the de-aliasing filter for the ADC that follows. The design of this filter depends on the system specifications and system design.

Spread LO generation - This block generates a spreaded LO signal; i.e. the LOI*PN and LOQ*PN signals. The input to the generation is fed by an oscillator which does not have any signal power at the frequency of the LO which is equal to the frequency of the RF signal.

Spreading sequence generation - This block generates PN (or So) from the oscillator.

Clock edge delay & correction - this block corrects the PN so that LO*PN*PN is equal to LO.

The Q-path in similar to the I-path except M1Q multiplies the RF signal by LOQ*PN.

In the example system given in Fig. 11, the calculation of the power is assumed to be done within the DSP unit. A correction signal is generated within the DSP. The method for correcting the error in the LO signal has been described in Fig. 10.

Sampling MUX Scheme

Fig. 12 depicts a new architecture that eliminates the problems associated with multiplying S_o with S_o. In this architecture, two branches of the structure shown in Fig. 12 are used. For each branch a different S_o is used. The two range from +A to –A, but are approximately 90 degrees out of phase; they are labeled

$S_{Qo}(t)$ and $S_{Io}(t)$; we shall assume A=1. The function of the MUX is to allow the output node to "see" the signal of only one branch at a time. During the period where $S_{Io}(t)S_{Io}(t-\tau)$ deviates from +1, the MUX allows the top branch to go through. Since $S_{Qo}(t)$ and $S_{Io}(t)$ are out of phase, during this cycle $S_{Qo}(t)S_{Qo}(t-\tau)$ equals +1. In the next cycle where $S_{Qo}(t)S_{Qo}(t-\tau)$ deviates from +1, the bottom branch, where $S_{Io}(t)S_{Io}(t-\tau)$ equals +1, goes through. In Fig. 13 we have plotted the values of $S_{Qo}(t)S_{Qo}(t-\tau)$ and $S_{Io}(t)S_{Io}(t-\tau)$ with time. By switch between the top and bottom branch in an appropriate manner, we completely eliminated the aliasing problems associated with equations (8) and (10). The clocking of the MUX can be derived from $S_{Qo}(t)$ and $S_{Io}(t)$. From Fig. 13 one can see that it is not necessary for $S_{Qo}(t)$ and $S_{Io}(t)$ to be exactly 90^o out of phase. That is, image rejection is effected without recourse to phase-shifting errors. The problem of accurate phase shifting has been re-casted as a time-domain technique. The minimum required phase difference is determined by the time period in which either $S_{Io}(t)S_{Io}(t-\tau)$ or $S_{Qo}(t)S_{Qo}(t-\tau)$ deviate from +1.

By introducing the two branches and the MUX, two non-ideal effects can degrade the performance of the structure. These are:

1. The switching delay in the MUX itself from one branch to the next.
2. A gain mismatch between the two branches.

In the following sections, the details of these two non-ideal effects will discussed as well as realistic solutions.

<u>Switching Delay</u>

The switching delay can cause aliasing in the an approximate manner the delay between $S_o(t)$ and $S_o(t-\tau)$ discussed earlier. However, this delay could be very small since the MUX switches at points were the signal at the top and bottom branches are equal (assuming no gain mismatching problem). Furthermore, by using clever circuit topologies, this non-ideal effect can be reduced significantly or removed all together.

To illustrate the effect of the switching delay we shall assume that after the MUX the signal looks as follows,

$$MUX_{out} = x(t)[1+\varepsilon_M(t,\tau)] \tag{17}$$

where $\varepsilon_M(t,\tau)$ has a period that is equal to twice the clocking rate of the MUX (denoted by T_{MUX} from here on) and takes the approximate form,

$$\varepsilon_M(t,\tau) = \begin{cases} \alpha(t-2nT_{MUX}) & 2nT_{MUX} < t < 2nT_{MUX}+\tau \\ 0 & 2nT_{MUX}+\tau < t < 2(n+1)T_{MUX} \end{cases} \tag{18}$$

where $\alpha(t-nT_{MUX})$ is a function that takes into account the switching delay between the two branches and n=0,1,2,3... . The amount of relative aliasing power reflected back to base band is given by the sum of the squares of the Fourier components of $\varepsilon_M(t,\tau)$ - i.e.

$$\text{aliased power} = \sum_{l=1,2\ldots} |\varepsilon_M(\omega_l)|^2 \tag{19}$$

where $\omega_l = 1/(2lT_{MUX})$. The amount of aliased power can be put in the form,

$$\text{aliased power} = \left[\frac{\tau}{2T_{MUX}}\right]^2 \sum_{l=1,2\ldots} \frac{1}{l^{2\xi}} |\eta_M(\omega_l)|^2 \tag{20}$$

where,

$$\varepsilon_M(\omega_l) = \left[\frac{\tau}{2T_{MUX}}\right]^2 \frac{1}{l^{2\xi}} |\eta_M(\omega_l)|^2 \tag{21}$$

and $|\eta_M(\omega_l)|^2$ is around unity and ξ is typically takes on value of 1 or 2, but both parameters are a function of α. As long as the sum in equation (20) converges fairly rapidly, the amount of aliasing power would be governed by the term $\left[\frac{\tau}{2T_{MUX}}\right]^2$. If we assume α=constant, $\xi = 1$, then,

$$|\eta_M(\omega_l)|^2 \sim \sin c^2 \frac{l\pi\tau}{2T_{MUX}} \tag{22}$$

One very simple method to help the sum converge faster is to place a first order filter before the MUX (in both branches) such that the pole is greater than but around $1/(2T_{MUX})$. Care must be made in setting this pole so that phase delay does not add to the wanted signal. Though the addition of the filter would help, if the MUX is built correctly, and ξ can be designed to equal 2, the sum in (19) converges relative fast.

Gain Mismatch

The second non-ideal effect is more important. If the gain of the top branch is G_T and the bottom is G_B, the output after the MUX can be described by,

$$\begin{aligned} MUX_{out} &= x(t)\frac{G_T + G_B}{2}[1 + \frac{G_T - G_B}{G_T + G_B} S_{MUX}(t)] \\ &= x(t)\overline{G}[1 + \delta S_{MUX}(t)] \end{aligned} \tag{23}$$

where $\overline{G} = (G_T + G_B)/2$, $\delta = (G_T - G_B)/(G_T + G_B)$, and S_{MUX} is the square wave [+1,-1] applied to the clock to switch from one branch to the next. The term $x(t)\delta S_{MUX}(t)$ represents aliasing of the wanted RF channel with frequency components located a harmonic of the frequency of S_{MUX}. If we assume SMUX has a duty cycle of 50%, the amount of relative aliasing powers is given by the equation

$$\text{aliasing power} = \frac{2\delta^2}{\pi^2} \sum_{n=1,2...} \frac{1}{n^2} \tag{24}$$

or,

$$\text{aliasing power (dB)} = 20\log\delta + 10\log\left[\frac{2}{\pi^2}\right] + 10\log\left[\sum_{n=1,2\ldots}\frac{1}{n^2}\right] = 20\log\delta - 7dB + 2.2dB \tag{25}$$

For a gain mismatch of 1%, the amount of aliasing power is equal to -44.8dB. Since the amount of mismatch is deterministic, one should be able to minimize it to levels that are tolerable. One method in which this can be accomplished is to minimizes the about of power of the MUX using a differential closed loop configuration. This is illustrated in Fig. 14. Fig. 14 shows a method that can be used to minimize the amount of aliasing by using a differential gain control between the top and bottom branch. The output after the MUX, is given by

$$MUX_{out} = x(t)\frac{G_T + G_B + (g_t - g_b)v}{2}[1 + \frac{G_T - G_B + (g_t + g_b)v}{G_T + G_B + (g_t - g_b)v}S_{MUX}(t)] \tag{26}$$

where v is the gain control voltage and g_t and g_b are the gain control coefficients of the top and bottom branch respectively. The power measured after the output low pass filter can be represented as the sum of two components; (I) the wanted power,

$$wanted_{out} = \overline{|x(t)|^2}\left|\frac{G_T + G_B + (g_t - g_b)v}{2}\right|^2 \tag{27}$$

(ii) un-wanted (aliased) power,

$$un-wanted_{out} = \overline{|S_{MUX}(t)x(t)|^2}\left|\frac{G_T - G_B + (g_t + g_b)v}{G_T + G_B + (g_t - g_b)v}\right|^2 \tag{28}$$

If $v = (G_B - G_T)/(g_t + g_b)$ the un-wanted power equals zero, and the wanted power is equal to,

$$wanted_{out} = \overline{|x(t)|^2}\left|\frac{G_T + G_B}{2}\right|^2\left|1 + \frac{(g_t - g_b)}{(g_t + g_b)}\frac{(G_B - G_T)}{(G_T + G_B)}\right|^2 \tag{29}$$

The wanted power is modified by a second order term in mismatch between the top and bottom branch. Because of this property, the wanted power remains approximately the same with changes in v, while the unwanted power changes significantly with changes in v. Therefor in the closed loop, the power of the unwanted signal can be reduced to zero by minimizing the power measured after

the output of the low pass filter. Note that the computation of the power and the minimization procedure can all be done in DSP. An important consideration in designing the loop response. The relaxation time constant of the loop has to be faster then the variation of the power of the wanted signal seen at the input antenna. Once the receiver has found the minimum point the system will response much faster to find a modified minimum point.

Example of RF receiver with the MUX Scheme

Fig. 15 illustrates a system diagram of the "MUX sampler" topology with the gain correction scheme. The items in the I-channel are as follows:

M1II - M1II is the input mixer which multiplies the RF signal with a wave form LOI*PNI where PNI is a function that varies from +A to -A. PNI is typically a NRZ square wave, or a NRZ pseudo-random digital signal. The design of this mixer depends on the system specifications and system design.

M1IQ - M1IQ is the input mixer which multiplies the RF signal with a waveform LOI*PNQ where PNQ is a function that varies from +A to -A. PNQ is typically a NRZ square wave, or a NRZ pseudo-random digital signal. PNQ and PNI are related to each other. The design of this mixer depends on the system specifications and system design.

M2II (M2IQ) - the second mixer multiplies the output of M1II (M1IQ) by PNI (PNQ). The design of this mixer depends on the specifications of the system. Though it has not been shown in the system diagram a filter could be placed between M1II and M2II.

LPF1II, LPFIQ - This is a low pass filter. It is used to reduce the amount of out of band power, which may cause the following elements to compress in gain or distort the wanted signal. The design of these LPF's depends on system design specifications.

AGCII, AGCIQ - These AGC's are used to control the differential gain between the two branches containing M1II and M1IQ. This is done to reduce the amount

of gain mismatch between the two branches. The design of these elements depend on the system specifications and design.

MUXI - The top and bottom branches are MUX'ed together. The signal that is used to MUX the two branches can be derived from PNI and PNQ.

SUM1I - The summing element is used to remove any DC offsets. Any know art can be used.

LPF2I - This filter provides further filtering of the base-band signal. The design of this filter depends on the system specifications and system design trade offs.

GAIN-I - These elements provide a significant amount of gain to the base band signal. The design of this gain unit depends on the system specifications and system design trade offs.

LPF3I - This filter is the de-aliasing filter for the ADC that follows. The design of this filter depends on the system specifications and system design.

Similar statements can be made for the items in the Q-channel.

The items in the synthesizer section are:

Spread LO generation - This block generates a spread LO signal; i.e. the LOI*PNI, LOQ*PNI, LOI*PNQ, and LOQ*PNQ signals. The input to the generation is fed by an oscillator which does not have any signal power at the frequency of the LO which is equal to the frequency of the RF signal.

Spreading sequence generation - This block generates PNI and PNQ from the oscillator.

Generation of MUX signal - this block generates the MUX'ing signal for MUXI and MUXQ.

In this example the amount of gain mismatch between the two branches (prior to the MUX's) is calculated within the DSP. The DSP also provides a correction signal to correct for the gain mismatch. The correction signals for the RF I-

signal are labeled +VI and –VI; +VI is for the top branch and –VI for the bottom branch. The same procedure is used for the RF Q-signal path.

IV. Summary of Inventions

- In a direct conversion receiver the local oscillator has a frequency equal to the RF carrier wave. This causes problems in terms of LO-RF leakage. In the scheme we have proposed, two well-defined functions are multiplied together within the RF signal path such that a virtual LO signal is multiplied with the RF signal. That is, $LO=(signal_1*signal_2)$ where $signal_1$ and $signal_2$ do not contain a significant amount of power at the RF carrier (see Fig. 16). If said signals generate too much power at the RF carrier they may degrade the signal to noise ratio of the wanted RF band. Either $signal_1$ or $signal_2$ can select the RF band. Up to now we have assumed $signal_1=LO*S_o$ and $signal_2=S_o$. However, *signal1* and *signal2* can be any signals that satisfy the conditions mentioned above. $Signal_1$ and $signal_2$ can be selected such that they help system performance and the integration of the RF system.
- Signal 1 and Signal 2 are not generated by modulating the said LO signal.
- There will always be an error in generating the LO signal from *signal1*signal2*; i.e. $signal_1*signal_2=LO+error(t)$. The error term can be minimized by using a closed feed back loop which encompasses taking a measurement which indicates the magnitude of the error term, while modifying a parameter that modifies this error term.
- Using a MUX and a parallel path for the RF signal, the delay error problem can be converted to a gain error problem.
- By adjusting the differential gain of the two branch before they are MUX'ed the gain mismatch error can be reduced to zero. This can be accomplished by measuring the amount of power at base band and minimizing it with respects to the differential gain applied between the two branches.

V. Generalization of Invention

The embodiments of the invention in which an exclusive property or privilege is claimed are defined as follows:

- several well-defined functions (labeled *signal1*, *signal2*, *signal3*, etc..) are multiplied together within the RF, IF, or/and base band signal path such they result is multiplying the said RF, IF, or/and base band signal by a LO signal; see Fig. 17. This results in either down converting or up converting the said RF, IF, or/and base band signal. The said LO signal takes the form, $LO=signal_1*signal_2*signal_3..*etc.$ The *signals* are constructured such that they do not contain a significant amount of power at the wanted RF frequency or any frequencies that degrade the performance of the RF receiver or transmitter. Any of the said *signals* can either select the RF band in a receiver topology, or set the RF frequency in a transmitter topology.
- There will always be an error in generating the LO signal from $signal1*signal2*signal_3..*etc$; i.e. $signal_1*signal_2*signal_3..*etc\ =LO+error(t)$. The error term can be minimized by using a closed loop configuration. This closed loop comprises of making a measurement that indicates the level of this error and than modifying a parameter which modifies this error term. The closed loop is designed such that the error term is minimized.
- The contain of said signals can vary with time in order to reduce the effects of the said errors.
- The said signals can in general be pseudo random or periodic functions of time.
- The said measurement can be the form of a power measurement.
- The said parameter can be the phase delay of either (or a combination of) *signal1*, *signal2*, *signal3*, ... etc.
- The said parameter can be the fall or rise times of either or a combination of *signal1*, *signal2*, *signal3*,... etc.
- The said parameter can be a combination of the said phase delay and the said fall or rise times.
- By using two parallel branches of the structure shown in Fig. 17 together with a input sampling MUX (see Fig. 18).

- The resulting error due to a gain mismatch between the two branches at the output of the said MUX can be minimized by adjusting the differential gain of the two branch before they are MUX'ed. This can be done by measuring the amount of power at base band and minimizing it with respects to the differential gain (see document for details).
- Filters may be placed between all the elements shown in Fig. 17.

All the above claims are stated within the document in greater detail.

Figure Captions

Figure 1: Illustration of the spread spectrum scheme as used in wireless CDMA systems.

Figure 2: The new down conversion scheme. $S_o(t)$ is a NRZ signal varying between +1 and -1.

Figure 3: (a) Example spectrum of $x(t)$. (b) Spectrum at point P_1 in Fig. 2. (c) Down converted spectrum at P_2.

Figure 4: The output spectrum of the second mixer in Fig. 2. The leakage terms have been included.

Figure 5: Illustration of $S_o(t)S_o(t-\tau)$.

Figure 6: Illustration of the aliasing terms due to $x(t)LO\varepsilon(t)$.

Figure 7: An illustration of $LPF(S_o)*LPF(S_o)$.

Figure 8: An example of generating of the $LO*S_o$ without generation power at the LO frequency. In this figure *PN* denotes S_o; i.e. $PN=S_o$.

Figure 9: An illustration of the error $\varepsilon_{LO}(t,\tau)$ under two different cases. Case I is due to a time delay in the S_o (or PN). Case II is because of an error is fall/rise between the true LO and the constructed LO.

Figure 10: A method to reduce aliasing power due to a delay error τ.

Figure 11: Fully system diagram of the timing corrected S_o system. Also included is the *I* and *Q* channels.

Figure 12: MUX sampling scheme.

Figure 13: Illustration of $S_{Qo}(t)S_{Qo}(t-\tau)$, $S_{Io}(t)S_{Io}(t-\tau)$, and S_{MUX}.

Figure 14: A method to reduce the amount of aliasing due to gain mismatch between the two arms.

Figure 15: Complete system diagram using the sampling MUX topology.

Figure 16: An illustration of the main claims of this patent.

Figure 17: (a) Generalizations of the invention. (b) Generalization of the sampling MUX topology.

CLAIMS:

1. A radio receiver comprising:

a local oscillator for generating a local oscillator signal having a frequency that is an integral multiple of the desired mixing frequency; and

a digital switch for mixing an incoming signal with said local oscillator signal at a corresponding integral multiple;

thereby mixing said incoming signal with said desired mixing frequency without leaking the desired local oscillator signal into the output.

2. A radio receiver comprising a timing correction circuit.

3. A radio receiver as claimed in claim 1 wherein said digital switch comprises a flip-flop and invertor sequence.

4. A radio receiver as claimed in claim 1 wherein the local oscillator frequency avoids the carrier frequency

5. A radio receiver as claimed in claim 1 further comprising an automatic feedback loop to match gain.

6. A radio receiver as claimed in claim 1 wherein said incoming signal is CDMA.

7. A radio receiver comprising a blind-eye sampler.

8. A radio receiver comprising an automatic feedback loop to align clock edges such that the desired signal characteristics match the measured signal characteristics.

9. An integrated circuit comprising the radio receiver of any one of claims 1 - 8.

10. A computer readable memory medium, storing computer software code in a hardware development language for fabrication of an integrated circuit comprising the radio receiver of any one of claims 1 - 8.

11. A computer data signal embodied in a carrier wave, said computer data signal comprising computer software code in a hardware development language for fabrication of an integrated circuit comprising the radio receiver of any one of claims 1 - 8.

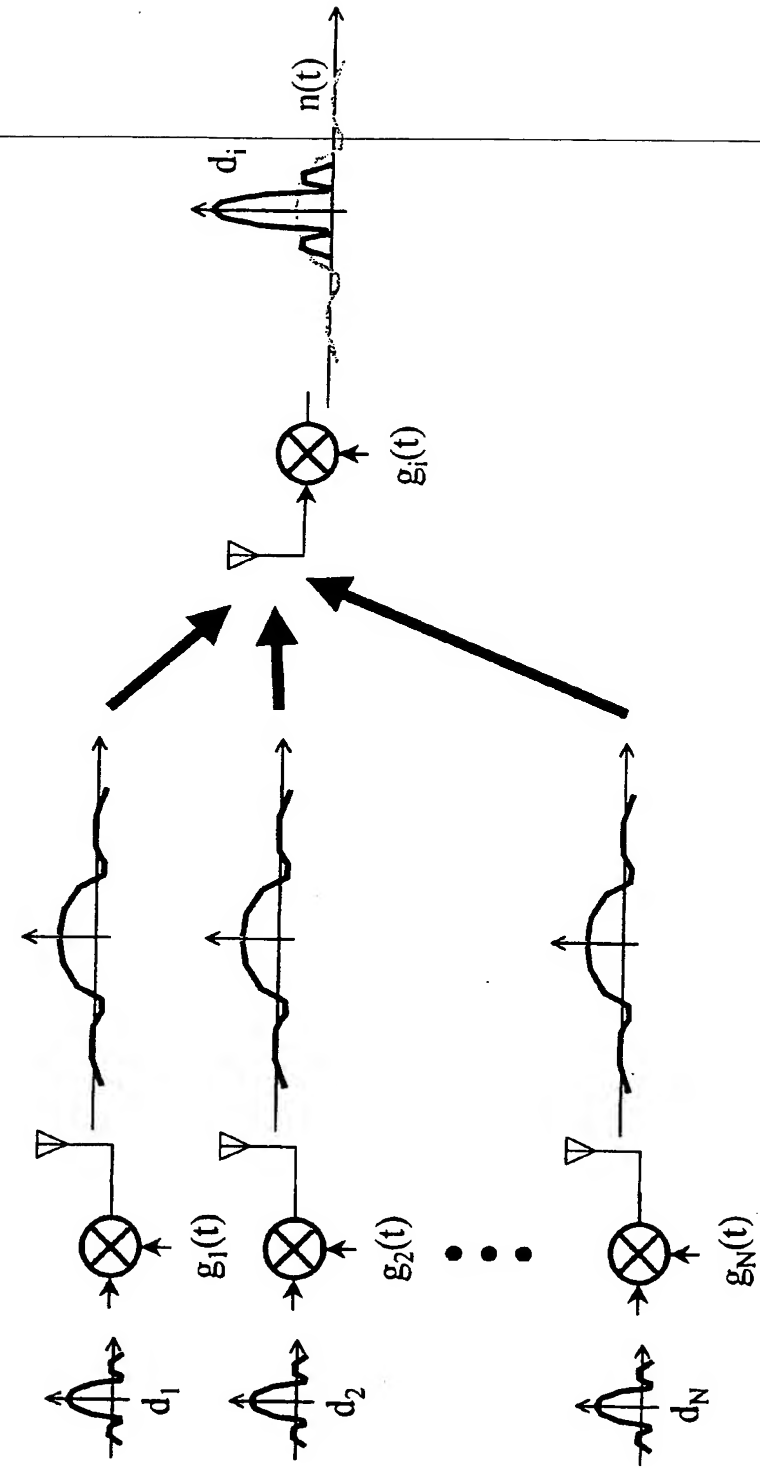


FIGURE 1

FIGURE 2

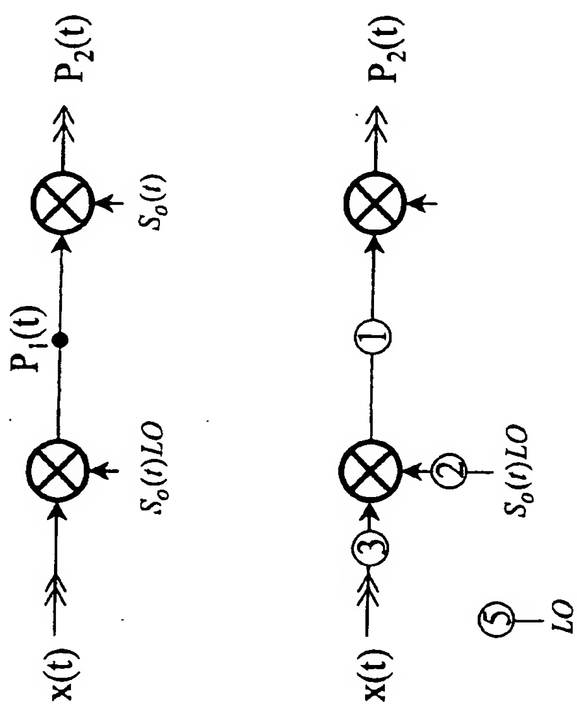
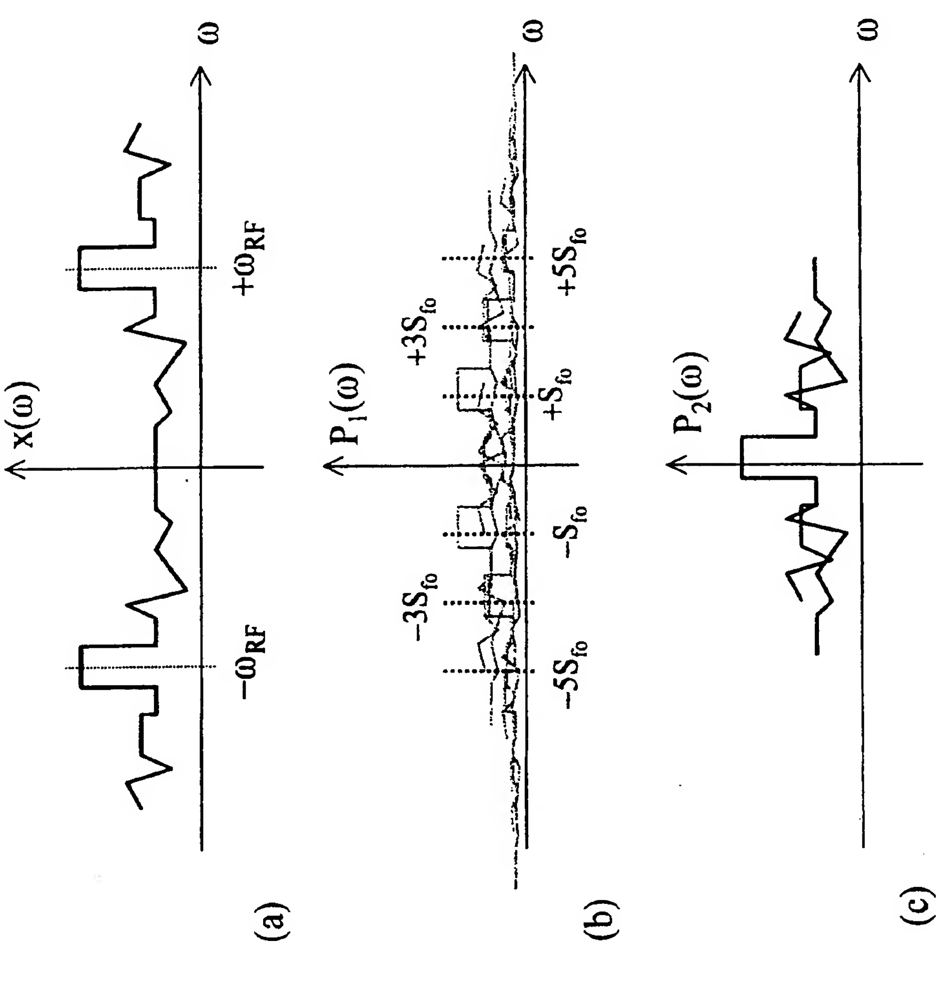


FIGURE 3



(a)
$X(\omega)$
$-\omega_{RF}$
$+\omega_{RF}$
ω
(b)
$P_1(\omega)$
$-5S_{fo}$
$-3S_{fo}$
$-S_{fo}$
$+S_{fo}$
$+3S_{fo}$
$+5S_{fo}$
ω
(c)
$P_2(\omega)$
ω

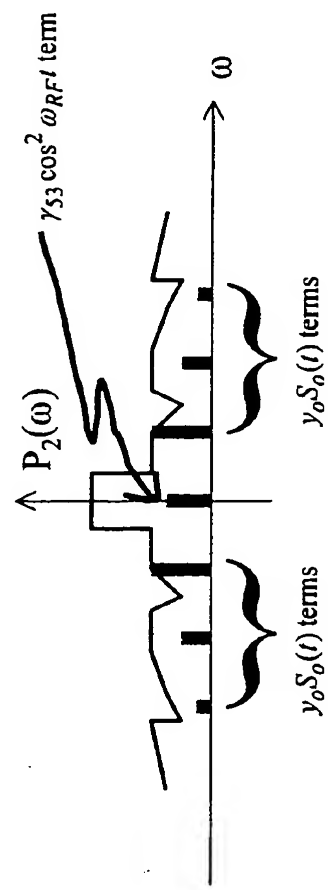

$\gamma_{53}\cos^2\omega_{RF}t$ term
$P_2(\omega)$
ω
$y_oS_o(t)$ terms
$y_oS_o(t)$ terms

FIGURE 4

FIGURE 5

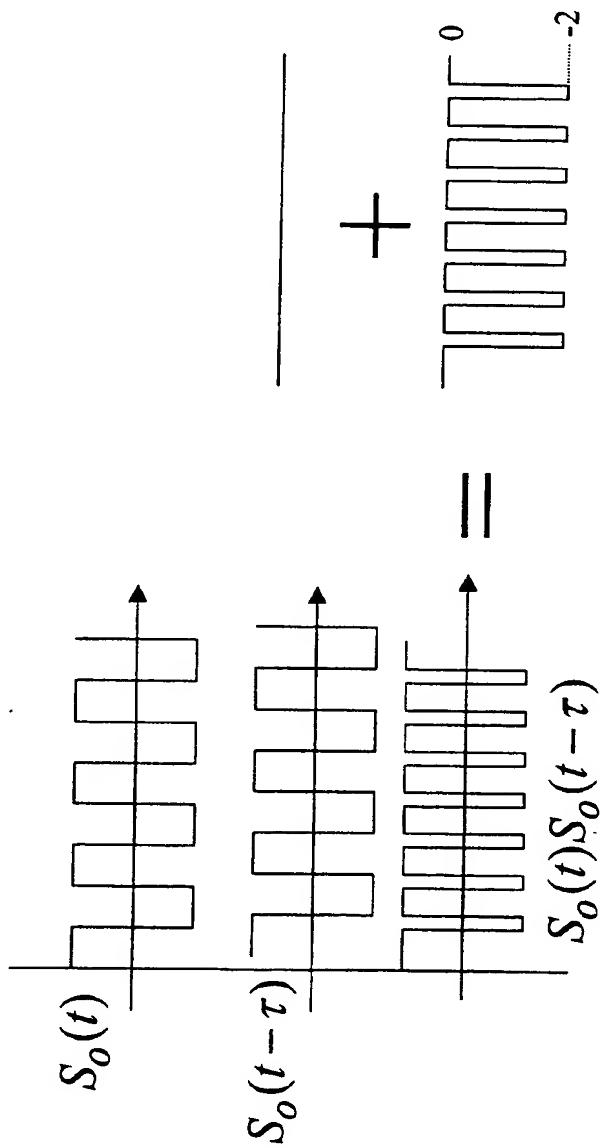
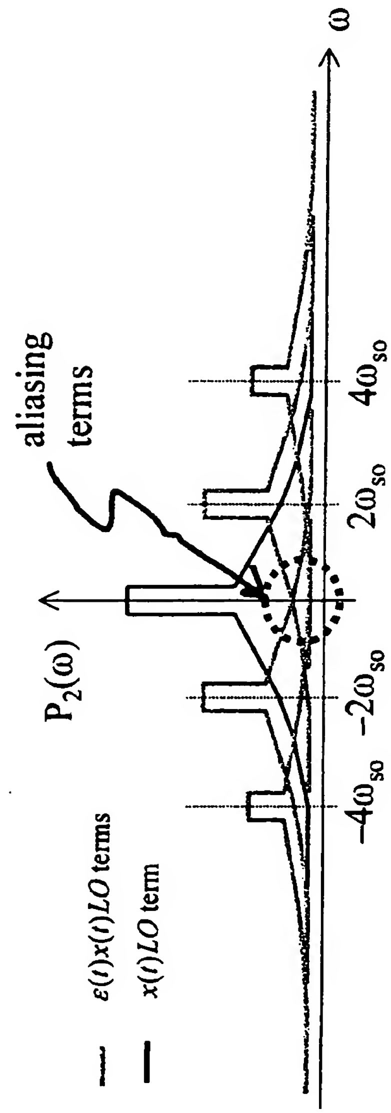
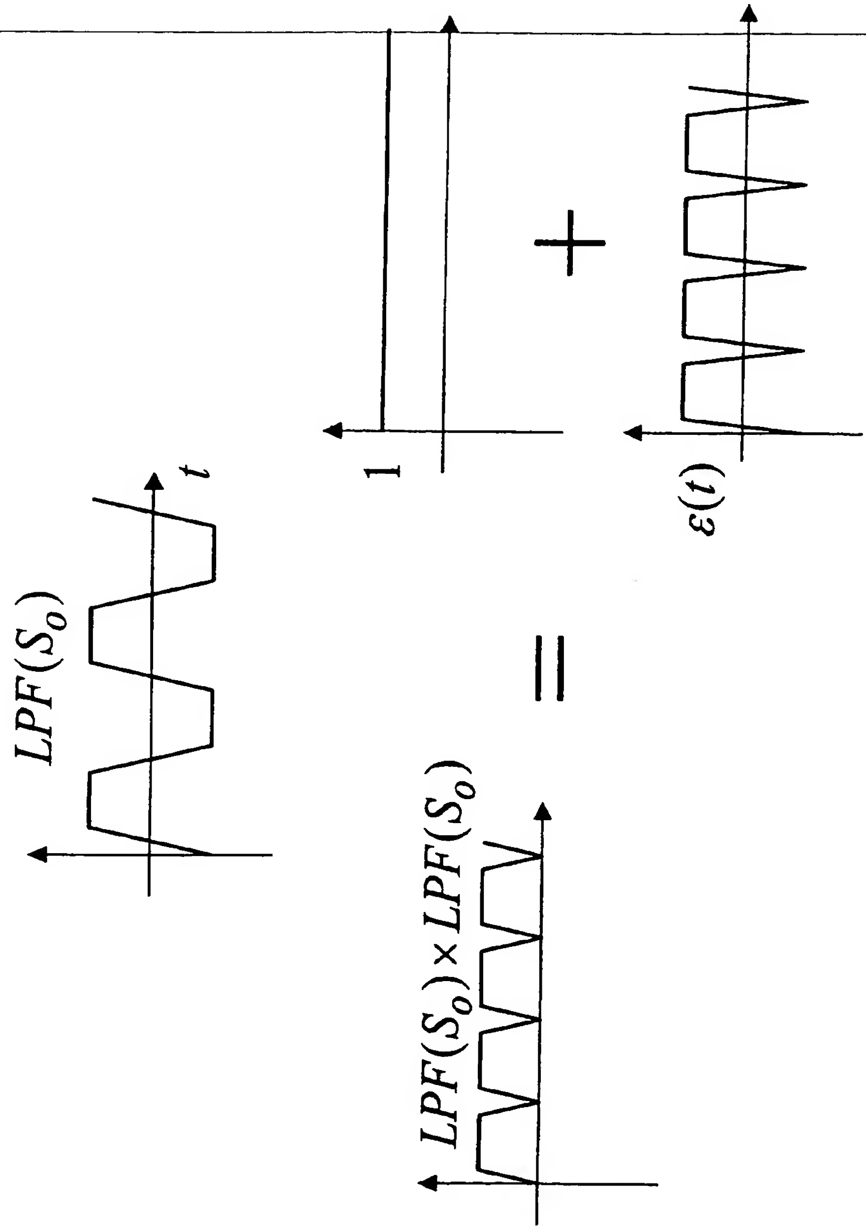


FIGURE 6



aliasing terms
$P_2(\omega)$
ω
$-4\omega_{so}$
$-2\omega_{so}$
$2\omega_{so}$
$4\omega_{so}$
$\varepsilon(t)x(t)LO$ terms
$x(t)LO$ term

FIGURE 7



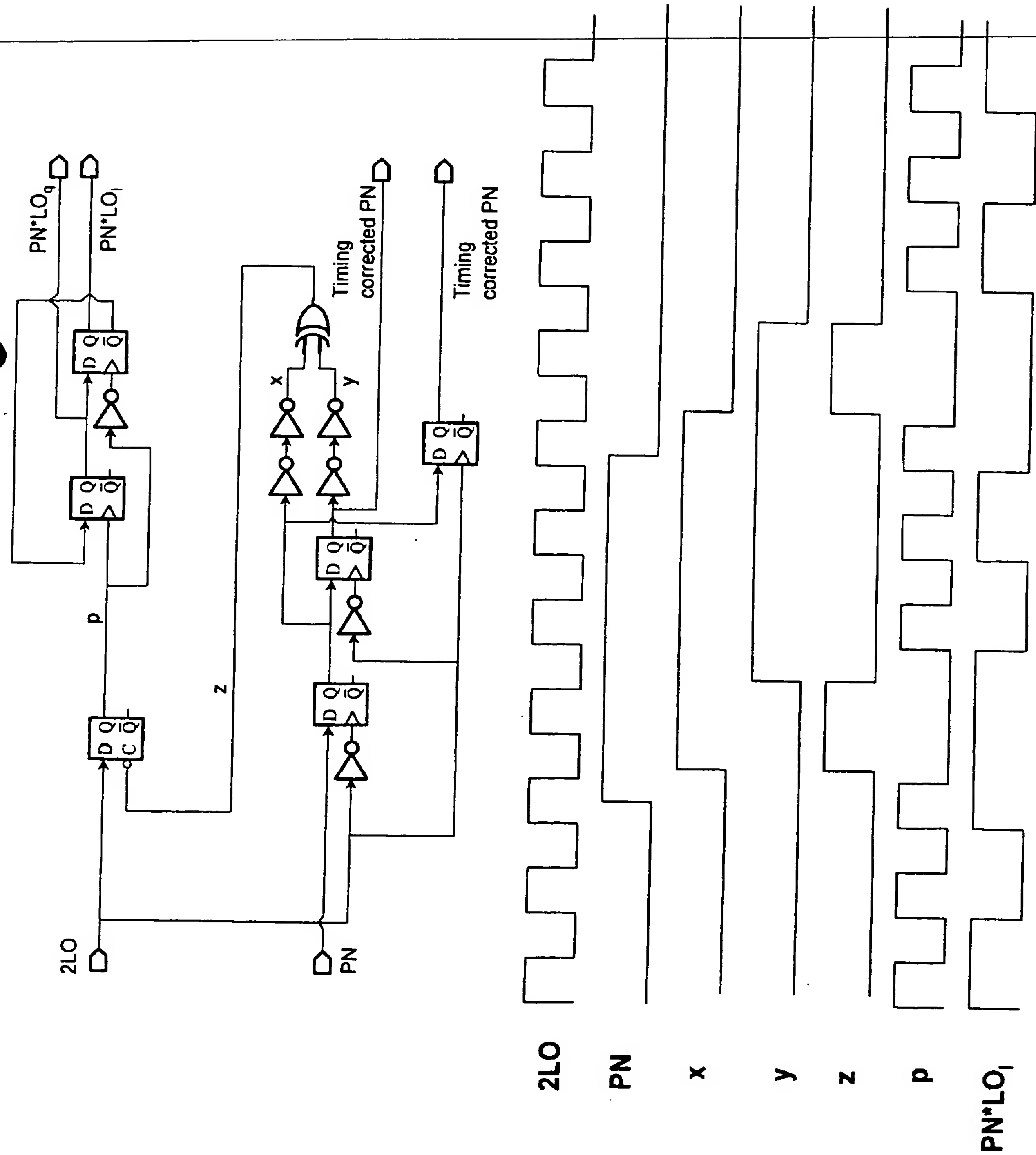
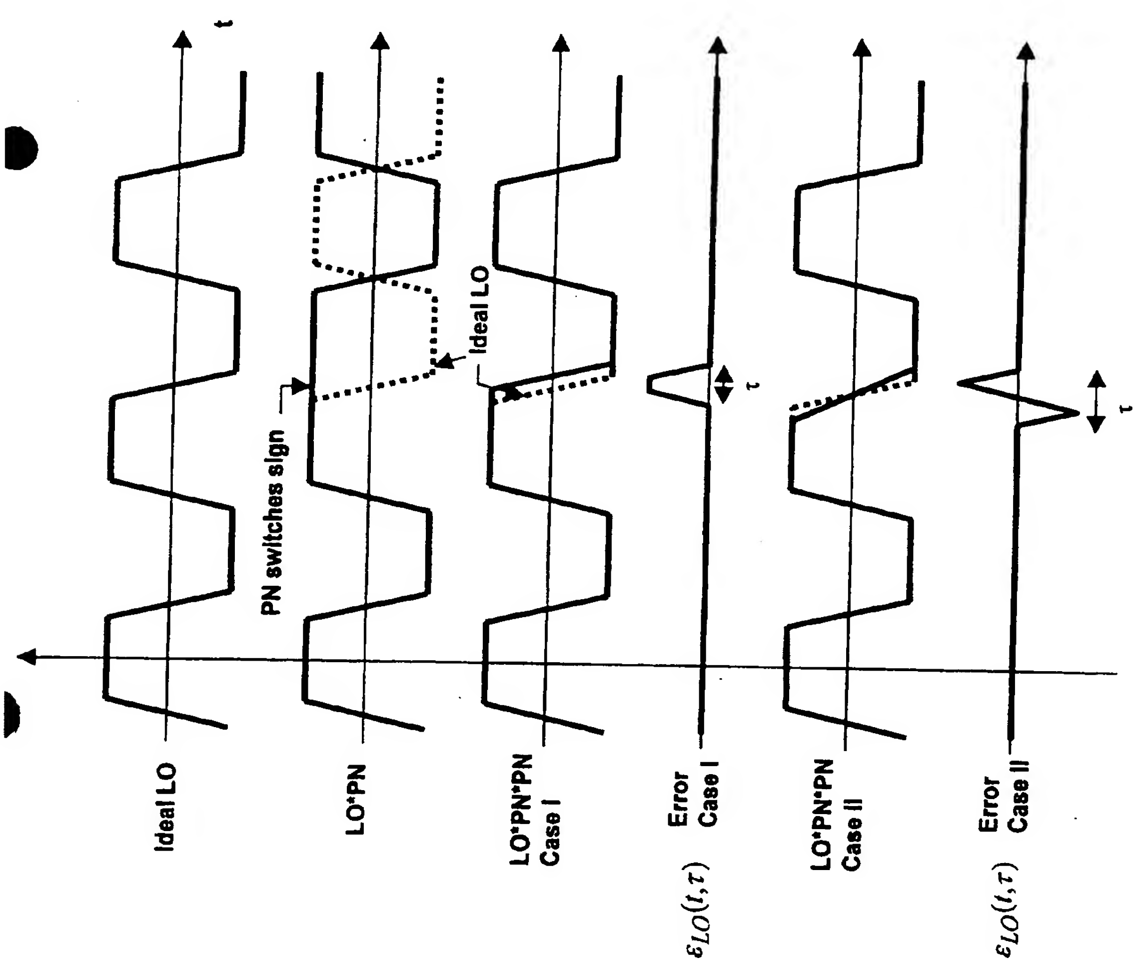

PN*LOq
PN*LOi
Timing corrected PN
Timing corrected PN
x
y
z
p
2LO
PN
2LO
PN
x
y
z
p
PN*LOi

FIGURE 8

FIGURE 9



Ideal LO
LO*PN
PN switches sign
Ideal LO
LO*PN*PN Case I
Error Case I
$\varepsilon_{LO}(t,\tau)$
LO*PN*PN Case II
Error Case II
$\varepsilon_{LO}(t,\tau)$
t
τ

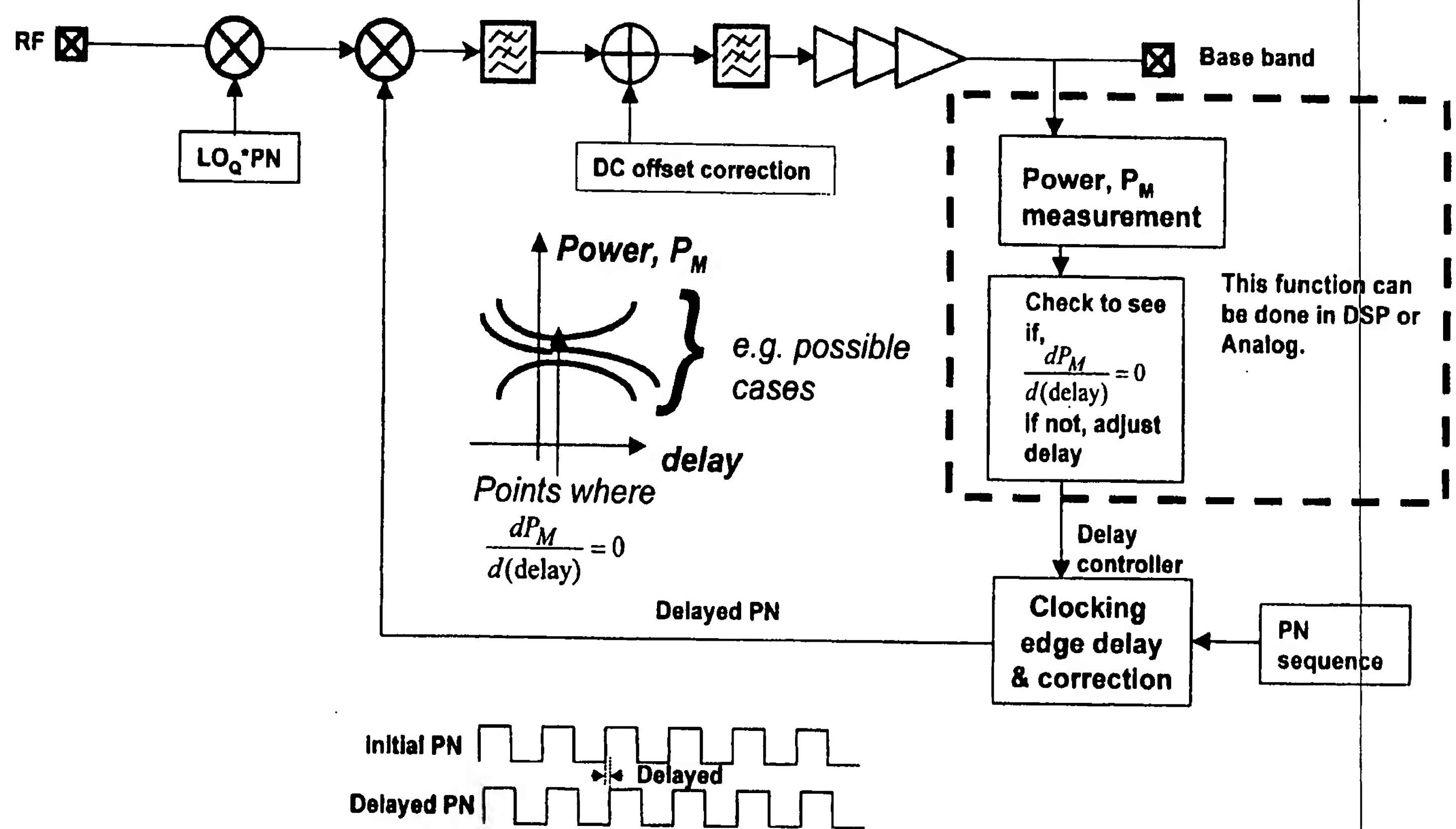

RF
LO_Q*PN
DC offset correction
Base band
Power, P_M measurement
Check to see if, $\frac{dP_M}{d(\text{delay})} = 0$ If not, adjust delay
This function can be done in DSP or Analog.
Power, P_M
e.g. possible cases
delay
Points where $\frac{dP_M}{d(\text{delay})} = 0$
Delay controller
Delayed PN
Clocking edge delay & correction
PN sequence
Initial PN
Delayed
Delayed PN

FIGURE 10

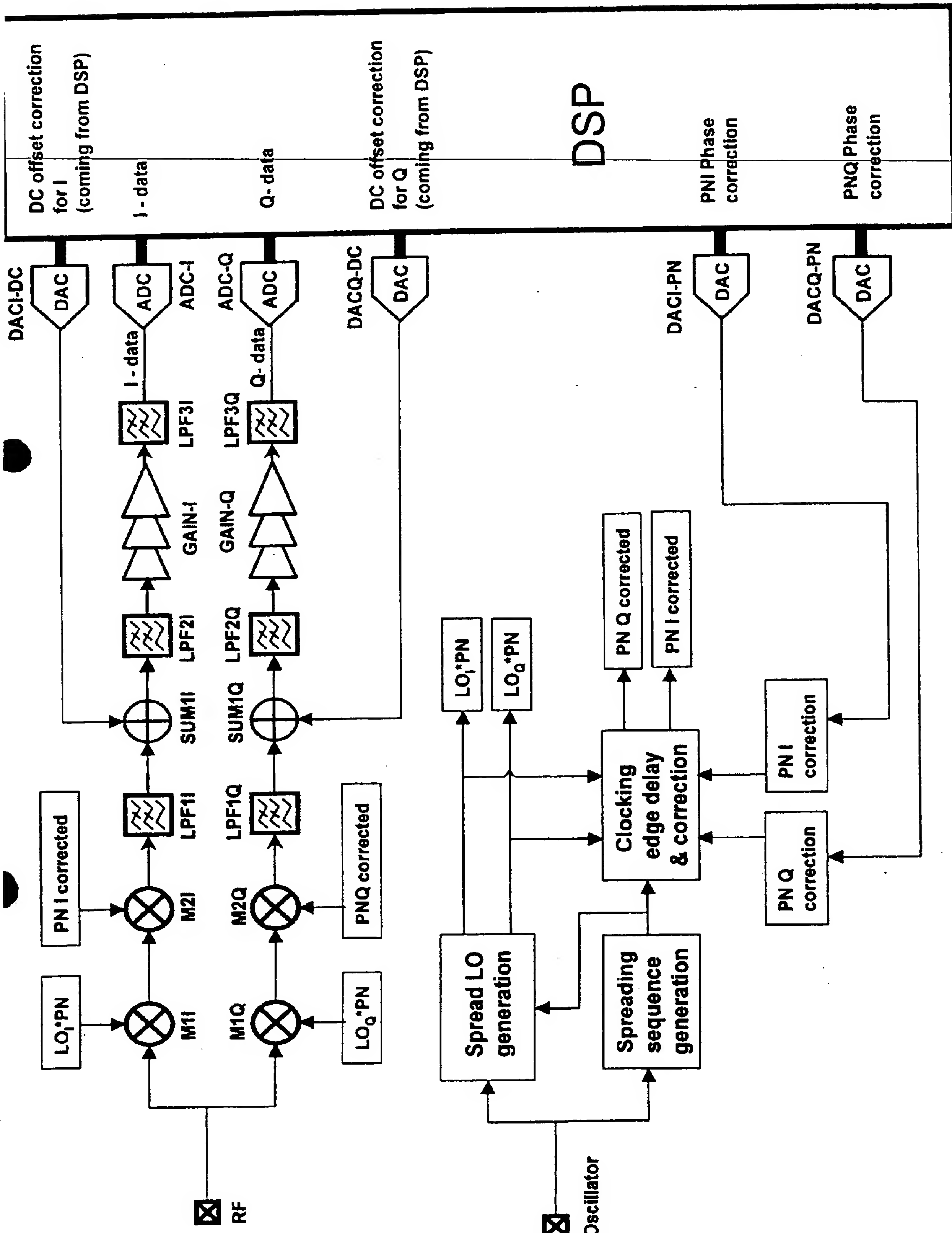
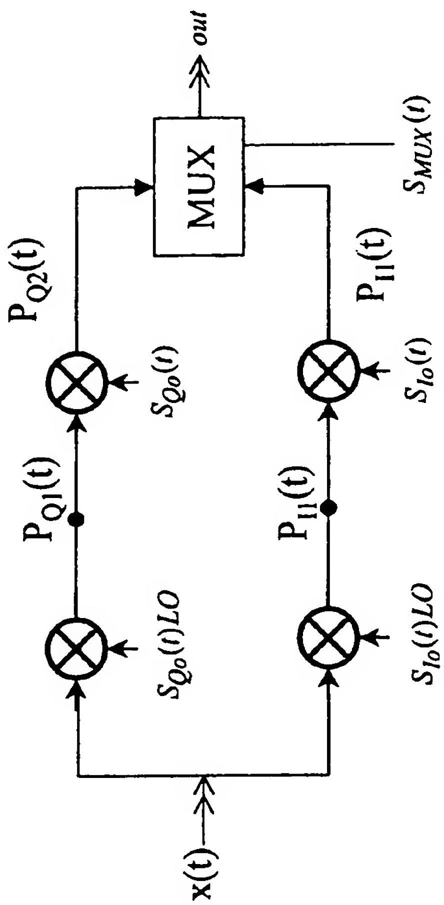
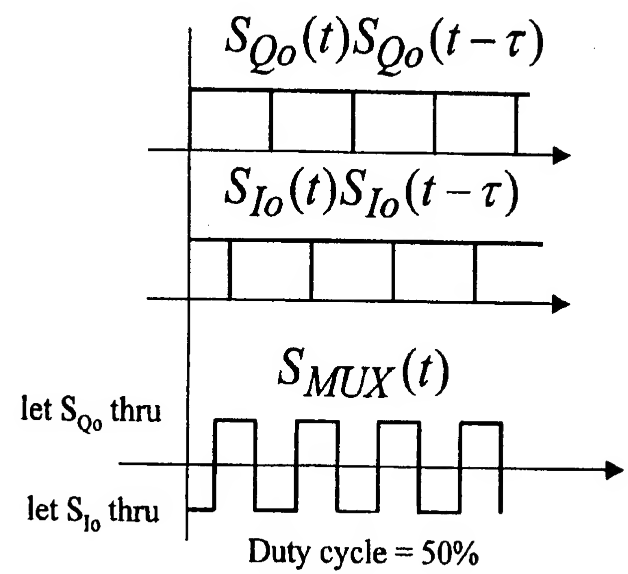

DSP
DC offset correction for I (coming from DSP)
I - data
Q- data
DC offset correction for Q (coming from DSP)
PNI Phase correction
PNQ Phase correction
DACI-DC
DAC
ADC
ADC-I
ADC-Q
ADC
DACQ-DC
DAC
DACI-PN
DAC
DACQ-PN
DAC
I - data
Q- data
LPF3I
LPF3Q
GAIN-I
GAIN-Q
LPF2I
LPF2Q
SUM1I
SUM1Q
LPF1I
LPF1Q
M2I
M2Q
M1I
M1Q
PN I corrected
PNQ corrected
LO$_I$*PN
LO$_Q$*PN
RF
LO$_I$*PN
LO$_Q$*PN
PN Q corrected
PN I corrected
Clocking edge delay & correction
PN I correction
PN Q correction
Spread LO generation
Spreading sequence generation
Oscillator

FIGURE 11

FIGURE 12

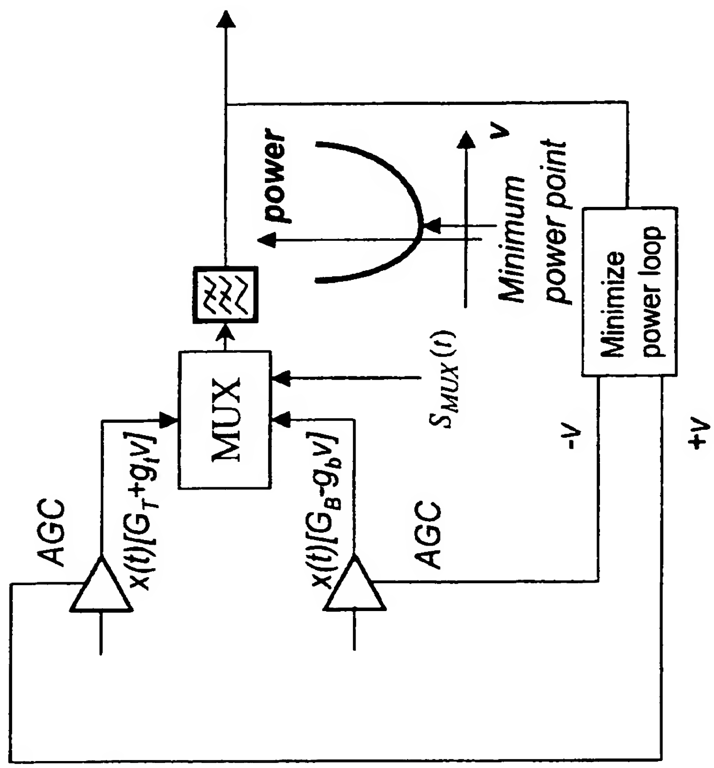




S_Qo(t)S_Qo(t − τ)
S_Io(t)S_Io(t − τ)
S_MUX(t)
let S_Qo thru
let S_Io thru
Duty cycle = 50%

FIGURE 13

FIGURE 14



power
v
Minimum
power point
Minimize
power loop
MUX
$S_{MUX}(t)$
-v
+v
AGC
$x(t)[G_T+g_t v]$
AGC
$x(t)[G_B-g_b v]$

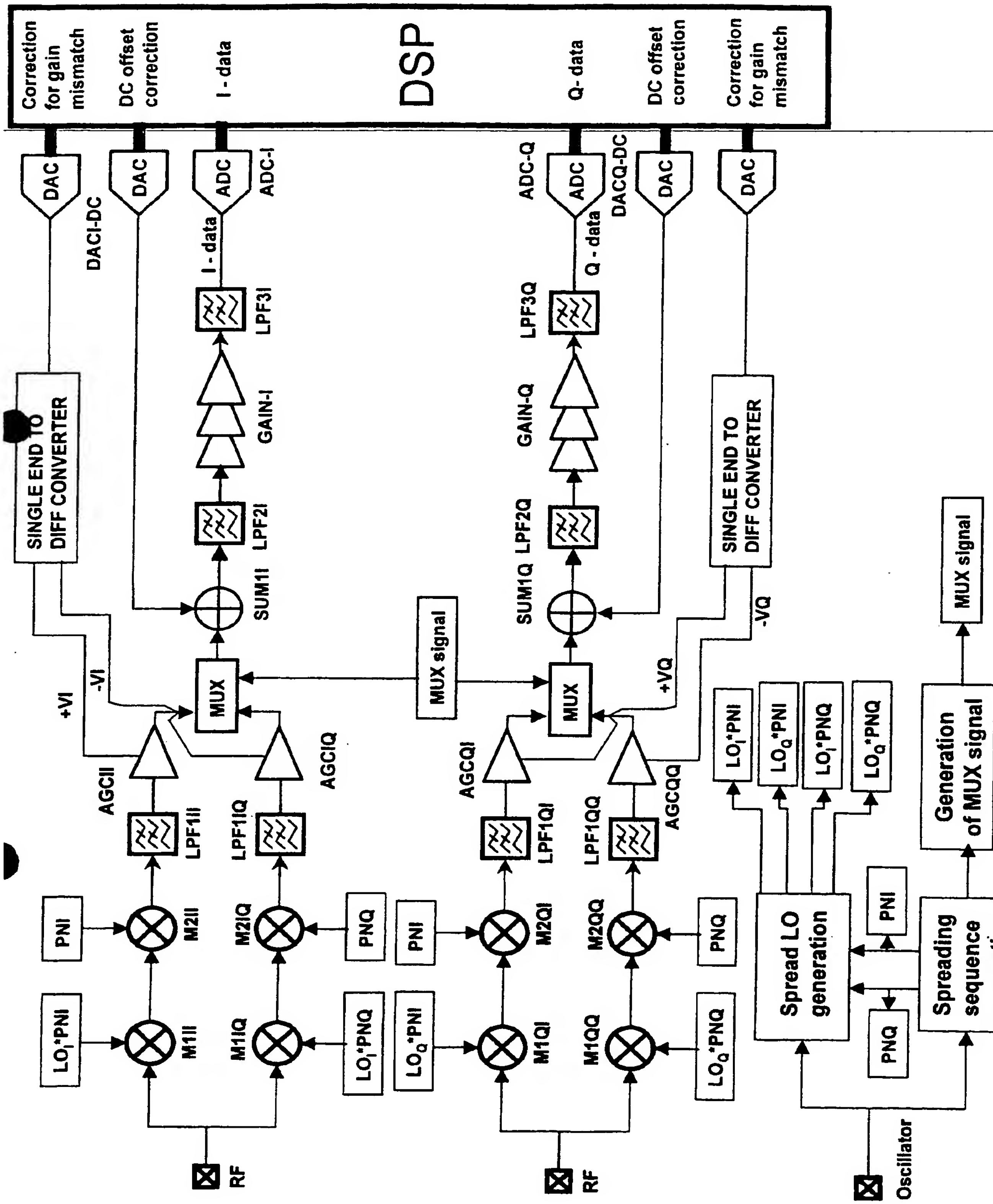
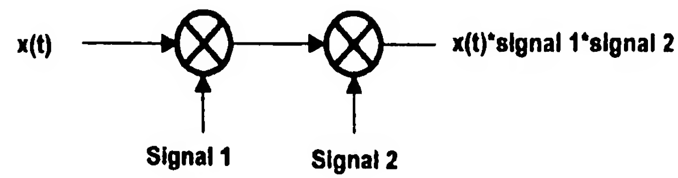


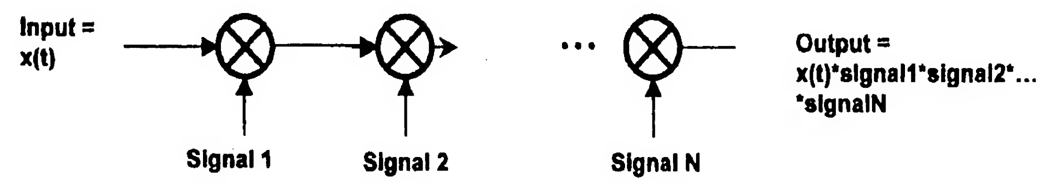
FIGURE 15

Gowling, Strathy & Henderson



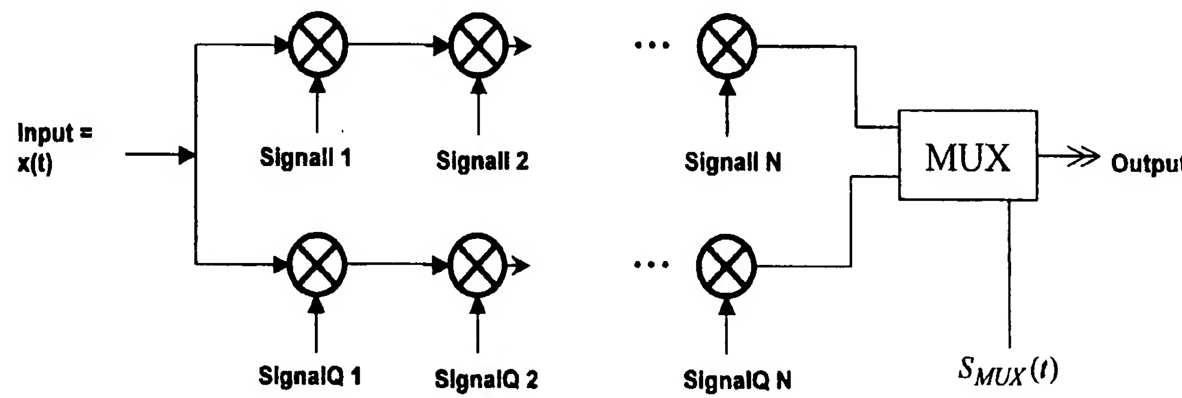
• *(Signal 1)*(Signal 2)=LO*
where LO has an oscillation frequency equal to the RF carrier wave or any frequency required in the RF system.
• Any errors in generating the required LO can be corrected by taking a measurement that changes with this error and incorporating it a closed loop to minimize the said error.
• The topology is not restricted to the receiver front end. It can be used in a RF transmitter where now x(t) is either the base band signal or a IF signal.
• *Signal 1 and Signal 2* both do not have any signal power (or a relatively small amount of power which can be tolerated by the system) at the RF carrier wave or at the intermediate frequency.
• *x(t)* is either the RF signal, IF signal, or a base band signal.
• either mixer can be implemented in digital or analog form.
• Signal1 and Signal2 signal content can change with time in order to reduce an errors in the systems.
• a filter or any other element can be placed between the first and second mixer.

FIGURE 16



• *x(t)* is either the RF signal, IF signal, or a base band signal.

(a)



(b)

FIGURE 17